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(71) Applicant *(for all designated States except US)*: ADAPTIVE CONTROL LIMITED [GB/GB]; Norwich Union House, 102-108 Above Bar, Southampton SO9 5HJ (GB).

(72) Inventors; and
(75) Inventors/Applicants *(for US only)* : ELLIOTT, Stephen, John [GB/GB]; 26 Elm Road, Winchester, Hampshire (GB). NELSON, Philip, Arthur [GB/GB]; 19 Chetwynd Drive, Southampton (GB). STOTHERS, Ian, McGregor [GB/GB]; 89 Newton Road, Eastleigh, Hampshire (GB).

(74) Agent: BOUTLAND, John, Anthony; 8 Heatherstone Avenue, Hythe, Southampton SO4 5LQ (GB).

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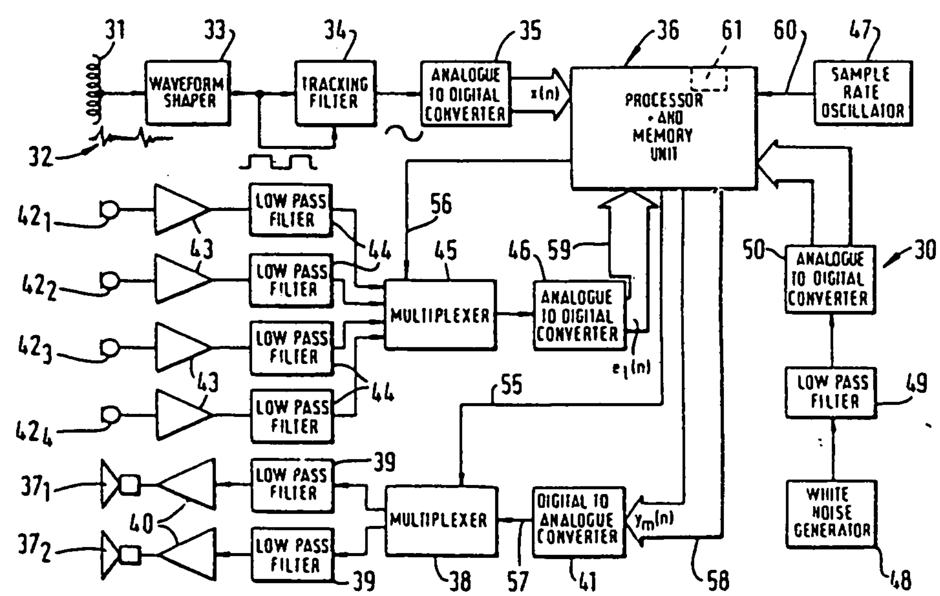

(57) Abstract

To reduce noise inside a motor car passenger compartment, two loudspeakers (37_1, 37_2) are driven by signals derived from a reference signal x(n) by adaptive filtering carried out by a programmed microprocessor and memory unit (36) which adapts the filtering in dependence on error signals $e_l(n)$ from four microphones (42_1, 42_2, 42_3 and 42_4) distributed in the passenger compartment. Reference filtering coefficients are initially determined by analysis of finite impulse responses when white noise is acoustically coupled from the loudspeakers (37) to the microphones (42), a white noise generator (48) being coupled to the unit (36). The reference signal x(n) is restricted to one or more selected harmonics or subharmonics of the fundamental noise frequency by a filter (34) which tracks the selected frequency. The selected frequency may be obtained from a coil (31) in the ignition circuit of the vehicle.

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Active vibration control.

BACKGROUND TO THE INVENTION

The invention relates to active vibration control.

As used herein, the term "vibration" includes sound or noise, and the invention is particularly concerned with active noise control.

In passenger compartments of cars (automobiles), where a significant noise component is harmonically related to the rotation frequency of a reciprocating engine used to drive the car, the sound levels at low frequencies in such enclosures are difficult to attenuate using conventional passive methods and can give rise to subjectively annoying "boom". A method of actively attenuating a simple sound field by introducing a single secondary sound source driven so that its output is in antiphase with the original ambient noise is described in general terms by B. Chaplin in "The Chartered Mechanical Engineer" of January 1983, at pages 41 to 47. Other discussions are to be found in an article entitled "Active Attenuation of Noise - The State of the Art" by Glenn E. Warnaka at pages 100 to 110 in Noise Control Engineering, May-June 1982 and in Internoise 83 Proceedings, pages 457 to 458 and 461 to 464, and Internoise 84 Proceedings, pages 483 to 488. Particular methods and apparatus are also described in British Patent Specification Nos. 1,577,322 and 2,149,614.

SUMMARY OF THE INVENTION

According to the present invention, an active vibration control system for reducing vibration generated by a primary source, is characterised in that at least one reference signal containing selected harmonics of the said primary source vibration is supplied to means driving a plurality of secondary vibration sources, such that vibration energy detected by sensor means operable to sense the vibration field established by the primary and secondary sources is reduced.

As used herein, the term "harmonic" includes "subharmonics".

Preferably the control system is operable in accordance with an algorithm which adjusts the outputs from the secondary sources so as to substantially reduce a cost function on a time scale comparable with the delays associated with the propagation of vibration from the secondary sources to the sensor means.

The present invention is particularly concerned with an active noise reduction system which can control the sound throughout an enclosure of a car, or at one or a number of "quiet zones" within it, and which can quickly adapt to changes in the excitation of the sound field due to changes in, for example, engine load or speed.

In order to ensure that the sound produced by the secondary vibration sources is of the same frequency as that produced by the engine, a signal related to the engine crankshaft rotation rate, for example a signal emitted by the engine ignition system, is used to generate a reference signal containing a number of sinusoids at harmonics (or subharmonics) of the engine crankshaft rotation frequency. These are known as engine order frequencies. These sinusoids may be obtained using a variety of methods outlined below. Rather than attempt to control all harmonics, only a selected set of engine order frequencies may, in accordance with the invention, be generated as the reference signal. For example, only the firing frequency (second engine order in a four cylinder car) and its second harmonic (fourth engine order) are used if the spectrum of the sound in the car is dominated by these components.

Alternatively, a signal containing all the engine order frequencies may be fed to a band pass filter which isolates only the particular frequency or frequencies exciting a particular resonance in the enclosure interior which could cause a "boom" to be excited. The advantage of reducing the number of frequencies fed to the filter is that an adaptive filter having fewer coefficients than

would otherwise be the case can be used. This makes implementation more efficient and allows faster adaption time. The faster adaption time is particularly important in automotive applications, for example, in which the active control system has to adapt sufficiently quickly to track changes in engine speed which may occur on a very short timescale.

Other reference signals may be obtained from transducers mounted on a road wheel hub or the suspension system of the car. Such reference signals would contain the harmonics of road wheel rotation or road noise. Transducers placed outside a vehicle may provide reference signals representative of wind noise. Reference signals from such sources may be broad band (random), in which case these reference signals may be fed directly to adaptive filters as described hereinafter. Alternatively, if the reference signals are periodic (deterministic), control at individual harmonics can be exercised as described hereinafter.

When the enclosed space is a motor vehicle interior, the secondary sources may be loudspeakers used as the low frequency drives of a car audio system.

Examples of methods in accordance with the invention for

generating the reference signals from a signal from an internal combustion engine are now discussed.

1. Selection of harmonics by filtering.

A signal is obtained from the primary vibration source which contains components at all harmonics prevalent in the sound in a vehicle powered by an engine. This signal is filtered so as only to leave the most important or dominant hermonics. Filtering is carried out by a filter whose centre frequency can be controlled by an external signal in such a manner that the critical filter frequencies have a constant ratio compared with the engine crankshaft rotation rate. This can be achieved by using, for example, charge coupled devices whose switching frequency is locked to the crankshaft rotation frequency, but can also be implemented as a program running on a microprocessor, as described hereinafter.

2. Selection of harmonics within a band by fixed filtering.

A primary source signal rich in harmonics is filtered by a band pass filter, having a centre frequency fixed at that of a pronounced "boom" in the car enclosure and a characteristic such that the reference signal only contains the harmonic(s) which are particularly exciting the boom. This may be extended such that the

filter contains a number of resonances at a number of boom frequencies of the car, or even such that the filter has a frequency response which models the acoustic response of the car interior to the primary excitation. The input signal to the filter may be a signal from the engine containing all important harmonics and may be in the form of a pulse train.

3. Generation of specific harmonics locked to the engine crankshaft rotation frequency.

This method may be accomplished by using phase lock loops to generate sinusoidal signals, with frequencies bearing an integer relationship to a square wave signal from the engine, which frequencies are then added together to form the reference signal. Alternatively, the signal derived from the engine can be used to control a number of tunable oscillators, each producing a sinusoid at a selected harmonic frequency. In one arrangement employing such a bank of tunable oscillators, the period of the square wave signal at the engine rotation rate is measured with a counter passed to a microprocessor which implements a number of digital oscillators using difference equations of the form

$$x_I(n) = \delta(n) + 2\cos(I\omega_o)x_I(n-1) - x_I(n-2)$$

where $\omega_o = 2\pi f_c/Nf_s$, I is the order of the harmonic or subharmonic

to be generated, f_c is the frequency o f the counter, which counts N pulses during a period, and f_s is the sample rate used for the difference equation. $\delta(n)$ is a unit sample sequence to initiate the oscillators formed by the difference equations above. The second and fourth harmonic may be generated for example.

It is advantageous if the sample frequency (f_s) is derived from the counter frequency (f_c) by a frequency division circuit for example, so that the ratio f_s/f_c is exactly an integer number.

An alternative difference equation which may be used to implement the digital oscillator has the form of a series approximation to a trigonometric function, for example :-

$$x_I(n) = 1 - \frac{y^2}{2!} + \frac{y^4}{4!} - \frac{y^6}{6!} + \frac{y^8}{8!} \simeq \cos(y)$$

The variable y is the accumulated phase of the oscillator, given by

$$y(n) = \sum_{i=0}^{n} \omega_i(n)$$

$\omega_i(n) = I\omega_o$ where ω_o is calculated from the measured period of the reference signal, as above, for every new sample (n). The series approximation above can be used for y(n) in the range $-\pi/2 < y(n) < \pi/2$. For values of y(n) outside this range the symmetry properties of the cosine waveform are utilised, until $y(n) > \pi$. When $y(n) \gg \pi$, the natural overflow properties of the two's complement number

representation is used to allow $y(n)$ to "wrap-around" to $y(n) \geqslant -\pi$ and the series expansions and symmetry properties discussed above are again used. In this way the digital representation of $y(n)$ is kept within the range $-\pi < y(n) < \pi$, and $x_I(n)$ is within the range ± 1.

The calculations of the coefficients used in the difference equations forming the digital oscillators, together with the difference equations themselves may be implemented on a dedicated processor, or may form part of the program which also implements a controller that generates the outputs used to drive the secondary sources from the reference signals described above.

The controller is designed to be adaptive so as to quickly track changes in engine speed and load. The outputs of the secondary sources are adaptively controlled so that some measurable cost function is minimised. This cost function would typically be the sum of the mean square outputs from a number of microphones in the enclosed space. The controller can be implemented as a digital adaptive FIR filter, using the basic update algorithm described by S. Elliott and P. Nelson in "Electronics Letters", at pp. 979-981, 1985. A number of additions must be made to this basic algorithm,

however, to enable it to work quickly and efficiently in this particular example. The basic algorithm, referred to hereinafter as the stochastic gradient algorithm, is presented below, in order to highlight the necessary alterations. If the i'th coefficient of the adaptive filter driving the m'th secondary source at sample number n is $w_{mi}(n)$, then each of these coefficients should be adjusted at every sample according to the equation

$$w_{mi}(n + 1) = w_{mi}(n) + \alpha \sum_{\ell=1}^{L} e_{\ell}(n) r_{\ell m}(n - 1)$$

where α is a convergence coefficient, $e_{\ell}(n)$ is the sampled output from the ℓ'th sensor and $r_{\ell m}(n)$ is a sequence formed by filtering the reference signal discussed above ($x(n)$, say) with a digital filter which models the response of the ℓ'th sensor to excitation of the m'th secondary source. These digital filters, generating each $r_{\ell m}(n)$, have only two coefficients in the implementation described in the "Electronics Letters" article, since control at only a single, fixed, frequency was being attempted. In the present example, however, the digital filters must model the relevant response over a <u>range</u> of frequencies, governed by the frequency range which the active system is attempting to control. It has been found that under certain circumstances a digital filter need only model the

overall delay in the response to ensure the stability of an adaptive filter. It is more common, however, to have the digital filters incorporate a delay and then some reverberant response. This may be implemented using either digital FIR or IIR filters whose coefficients are adjusted adaptively during an initialisation phase, so as to accurately match the desired responses. It is also possible to continue this initial adaption process during the operation of the active control system by feeding training signals to each secondary source which are suitably uncorrelated with each other and with the primary excitation. This may be necessary to track changes in the acoustic response of the enclosure. Alternatively, if the change is due to some well-defined cause, such as a passenger sitting down or a car window being opened, this change may be detected with mechanical transducers and the information used to switch between a cariety of filters modelling the response of the enclosure under a variety of conditions.

Another important consideration concerning the use of adaptive algorithm in this example concerns the effect of unwanted, low level, harmonic or subharmonic components in the reference signal. Suppose that the method used to generate the reference signal, as described

above, is designed to produce only I frequencies. Even though there may only be I desired harmonics in such a system, there will in practice also be a number of other harmonics or subharmonic frequencies at low level, because of the finite cut-off rate of the filters, for example. These components may also be being generated by the primary source and therefore present in the enclosure and hence at the outputs of sensors such as microphones and thus the adaptive algorithm will attempt to cancel them by enormously amplifying the low level, spurious harmonic reference signals. This can cause numerical overflow problems in the adaptive filter coefficients. This may be prevented in a number of ways:

(1) The use of only 2.I coefficients in the adaptive filter.

(2) The deliberate injection of random noise into the reference signal.

(3) The insertion of a "leak" into the algorithm so that, in the equation above, the past coefficient value, $w_{mi}(n)$, is multiplied by a factor close to but not equal to unity before being updated.

(4) The addition of an extra term in the update equation which minimises a cost function involving "effort" as well as

"error", as described in ISVR Technical Report No. 136, 1985, published by the Institute of Sound and Vibration Research, University of Southampton.

A number of other adaptive algorithms may also be implemented to adjust the coefficients of the digital filters in the controller driving the secondary sources. These alternative algorithms are best described in matrix form.

Assuming the availability of a sampled reference signal, $x(n)$, which is correlated with the output of the primary source, but is unaffected by the action of the secondary sources. The output to the m'th secondary sources, $y_m(n)$, may be obtained by passing this reference signal through a digital filter whose i'th coefficient is $w_{mi}(n)$ at the n'th sample, so that

$$y_m(n) = \sum_{i=0}^{I-1} w_{mi}(n)x(n-i)$$

The sampled output from the ℓ'th error sensor, $e_\ell(n)$, is equal to the sum of the contributions from the primary source, $d_\ell(n)$, and each of the secondary sources. The response of the path between the m'th secondary source and ℓ'th error sensor is modelled as a J'th order FIR filter with coefficients $c_{\ell mj}$ so that

$$e_\ell(n) = d_\ell(n) + \sum_{m=1}^{M} \sum_{j=0}^{J-1} c_{\ell mj} y_m(n-j)$$

Therefore

$$e_\ell(n) = d_\ell(n) + \sum_{m=1}^{M} \sum_{j=0}^{J-1} c_{\ell mj} \sum_{i=0}^{I-1} w_{mi}(n-j)x(n-i-j)$$

In order to obtain a matrix expression for the error surface we must now make the assumption that the filter coefficients in the controller are time invariant, i.e. the controller only adapts very slowly compared to the time scale of the response of the system to be controlled. Then $w_{mi}(n-j) = w_{mi}$ and

$$e_\ell(n) = d_\ell(n) + \sum_{m=1}^{M} \sum_{j=0}^{J-1} \sum_{i=0}^{I-1} c_{\ell mj} \; w_{mi} \; x(n-i-j)$$

If we let

$$r_{\ell m}(n) = \sum_{j=0}^{J-1} c_{\ell mj} x(n-j)$$

where the sequences $r_{\ell m}(n)$ for each ℓ and m are called the filtered reference signals, then

$$e_\ell(n) = d_\ell(n) + \sum_{m=1}^{M} \sum_{i=0}^{I-1} w_{mi} r_{\ell m}(n-i)$$

or $\quad e_\ell(n) = d_\ell(n) + \underline{r}_\ell^T(n)\underline{w}$

where

$$\underline{r}_\ell^T(n) = \Big[r_{\ell 1}(n) \quad r_{\ell 1}(n-1) \; \ldots \; r_{\ell 1}(n-I+1) \quad r_{\ell 2}(n) \; \ldots \; r_{\ell 2}(n-I+1) \quad r_{\ell 3}(n) \; \ldots \quad r_{\ell M}(n) \; \ldots \; r_{\ell M}(n-I+1) \Big]$$

$$\underline{w}^T = \Big[w_{10} \quad w_{11} \quad w_{1I-1} \quad w_{20} \quad w_{2I-1} \quad w_{30} \; \ldots \quad w_{M0} \quad w_{MI-1} \Big]$$

So if

$$\underline{e}^T(n) = \Big[e_1(n) \quad e_2(n) \quad \ldots \quad e_L(n) \Big]$$

$$\underline{d}^T(n) = \Big[d_1(n) \quad d_2(n) \quad \ldots \quad d_L(n) \Big]$$

then

$$\underline{e}(n) = \underline{d}(n) + \underline{R}(n)\underline{w}$$

where

$$\underline{R}^T(n) = \left[\underline{r}_1(n) \quad \underline{r}_2(n) \quad \ldots \quad \underline{r}_L(n)\right]$$

If the cost function is written as

$$J = E\left(\sum_{\ell=1}^{L} e_\ell^2(n)\right) = E\left[\underline{e}^T(n)\ \underline{e}(n)\right]$$

where E is the expectation operator, then

$$J = E(\underline{d}^T(n)\underline{d}(n)) + 2\underline{w}^T E(\underline{R}^T(n)\underline{d}(n)) + \underline{w}^T E(\underline{R}^T(n)\underline{R}(n))\underline{w}$$

Using the standard theory of matrix quadratic forms, the minimum of J, J_{min}, is obtained with

$$\underline{w}_{opt} = -\underline{A}^{-1}\underline{b} = -E(\underline{R}^T(n)\underline{R}(n))^{-1}E(\underline{R}^T(n)\underline{d}(n))$$

giving

$$J_{min} = c - \underline{b}^T\underline{A}^{-1}\underline{b}$$

$$= E(\underline{d}^T(n)\underline{d}(n)) - E(\underline{d}^T(n)\underline{R}(n))E(\underline{R}^T(n)\underline{R}(n))^{-1}E(\underline{R}^T(n)\underline{d}(n))$$

The true gradient may be written

$$\frac{\partial J}{\partial \underline{w}} = 2E(\underline{R}^T(n)\underline{d}(n) + \underline{R}^T(n)\ \underline{R}(n)\ \underline{w}(n))$$

$$\therefore \quad \frac{\partial J}{\partial \underline{w}} = 2E(\underline{R}^T(n)\ \underline{e}(n))$$

The time domain steepest descent algorithm may thus be written as

$$\underline{w}_{k+1} = \underline{w}_k - 2\mu E(\underline{R}^T(n)\ \underline{e}(n))$$

In a practical implementation the true expectation could be

approximated by an MA or AR averaging process. Alternatively, the instantaneous gradient could be used to update each filter coefficient every sample, as in the time domain "stochastic gradient" algorithm

$$\underline{w}(n+1) = \underline{w}(n) - 2\mu \underline{R}^T(n)\ \underline{e}(n),$$

which is the matrix representation of the algorithm described above. It is clear from this formulation that the stability and convergence properties of this algorithm, in the limit of slow adaption, are governed by the eigenvalue spread of the matrix $E(\underline{R}^T(n)\ \underline{R}(n))$. This matrix depends only on the response of the system to be controlled, the positioning of the sources and sensors within that system and the spectral properties of the reference signal, $x(n)$. It is possible that an unfortunate placing of these sources and sensors could cause this matrix to become ill conditioned, so that it has a large eigenvalue spread. This would create very slow "modes" in the convergence properties of such an algorithm.

This problem could be removed by using a Newton's method algorithm, the exact form of which may be written as

$$\underline{w}_{k+1} = \underline{w}_k - 2\mu E([\underline{R}^T(n)\ \underline{R}(n)])^{-1} E(\underline{R}^T(n)\ \underline{e}(n))$$

Again, various types of averaging could be used to give a practical approximation to the expectation operator. It should be

noted however that the time-independent matrix $E(\underline{R}^T(n)\ \underline{R}(n))$ depends only on the response of the system to be controlled and on the reference signal, and these are assumed to be known and stationary. This suggests a variety of stochastic Newton's method ("SNM") algorithms. The most obvious of these uses a modified or "normalised" set of reference signals $\underline{Q}(n)$, such that

$$E([\underline{R}^T(n)\ \underline{R}(n)])^{-1}\ \underline{R}^T(n) = \underline{Q}^T(n).$$

The computation of each of these reference signals will take somewhat longer than for $\underline{R}(n)$ alone since none of the elements of $\underline{Q}(n)$ are necessarily time delayed versions of any other elements. The complete SNM algorithm, again using instantaneous versions of $E(\underline{R}^T(n)\ \underline{e}(n))$, becomes

$$\underline{w}_{n+1} = \underline{w}_n - 2\mu \underline{Q}^T(n)\ \underline{e}(n).$$

BRIEF DESCRIPTION OF THE DRAWINGS

Embodiments of the invention will now be described, by way of example only, with reference to the accompanying drawings, wherein :-

Figure 1 is a block schematic diagram of an active noise control system associated with an enclosed space,

Figures 2(a), (b) and (c) are graphical representations of the behaviour of an element of the system,

Figure 3 is a block diagram of one form of reference signal generator,

Figure 4 is a block diagram of another form of reference signal generator,

Figure 5 is a block diagram of a circuit which incorporates a microprocessor,

Figure 6 is a block diagram illustrating how two reference signals are combined,

Figure 7 is a flow chart illustrating processing in a particular embodiment of the invention,

Figure 8 is a schematic diagram illustrating a heterodyne and averaging method of obtaining inphase and quadrature components of an error sequence in another embodiment of the invention,

Figure 9 illustrates application of the invention to non-noise vibration control, and

Figure 10 illustrates a modification of the arrangement shown in Figure 9.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

In Figure 1 an enclosure 10, which is the interior of the passenger or driver compartment of an internal combustion engine driven vehicle, in this example, a motor car 100, is represented schematically together with an active sound control system 1 according to the invention. In this example, the system 1 employs two secondary sound sources 11, comprising two low frequency loudspeakers of a stereo audio system fitted to the car, and three acoustic sensors, comprising microphones 12. The loudspeakers 11 are driven by a controller circuit 13 which comprises a pair of adaptive filters 14. Each adaptive filter 14 drives a respective one of the loudspeakers 11 with an output signal 3 which the filter 14 produces as a result of its action on a reference signal 4 supplied thereto by a reference signal generator 15. The reference signal 4 is generated by the generator 15 from an input signal 16 which is periodic at the crankshaft rotation rate of the internal combustion engine 2.

The signal generator 15 may comprise a tracking filter.

The purpose of the outputs from the loudspeakers 11 driven by the controller 13 is to reduce the sound vibration field established by the primary and secondary sources, experienced within the enclosure 10. Since the primary source (engine 2) of the noise to be reduced is periodic, the reference signal 4 generated by the generator 15 is, in accordance with the invention, arranged to contain one or more sinusoidal components at harmonics (or subharmonics) of the crankshaft

rotation rate of the engine 2. The adaptive filters 14 are adjusted automatically by output signals 5 from the sensor-microphones 12, corresponding adjustment being made simultaneously to the outputs of the loudspeakers 11 so as to substantially minimise a cost function on a time scale comparable with the delays associated with the propagation of sound vibrations from the loudspeakers 11 to the microphones 12. The cost function may comprise the sum of the mean square outputs of the microphones 12.

A decision is made beforehand as to what harmonics are to be selected; a decision which may vary from car to car.

The control system 1 does not employ a stored solution. Instead, it makes use of a plurality of closed loops, each loop comprising a microphone 12, the controller 13, and a loudspeaker 11, whereby signals from the microphone 12 are used to adapt the filters 14 controlling the loudspeaker 11, which has an influence on the output of the microphone as a result of acoustic response within the enclosure 10.

Each loop is accounted for by part of an algorithm which adjusts the outputs of the loudspeakers 11 as aforesaid, the algorithm being of the form :-

$$\underline{w}(n+1) = \underline{w}(n) - 2\mu\, \underline{R}^{T}(n)\, \underline{e}(n).$$

It will be noted that the loudspeakers 11 and the microphones 12 are distributed in the enclosure 10 in spaced relationship. The distribution, which varies from car to car, is adjusted to get substantial sound reductions throughout the enclosure 10.

It will also be noted that the system 1 employs as many closed

loops as the number of sensors (12) multiplied by the number of secondary sources (11).

The manner in which each secondary source 11 effects every one of the sensors 12 is reflected in the algorithms referred to herein.

Furthermore, the system 1 employs more sensors (12) than secondary sources (11), whereby a controlled reduction of primary source vibration is achieved. This contrasts with presently-known systems employing the same number of secondary sources as sensors, whereby near perfect cancellation can be achieved at the sensor locations but vibration levels away from these locations are increased.

Since the reference signal 4 contains harmonics of the input signal 16, which signal is periodic at the engine crankshaft rotation rate, the reference signal 4 contains engine order frequencies. The signal generator 15 is arranged to select engine order frequencies that ensure that the sound produced by the loudspeakers 11 is of the same frequency or frequencies as the sound produced in the enclosure 10 by the engine 2, even during changes in engine conditions such as load or speed. The number of engine order frequencies in the reference signal 4 is restricted so that the adaptive filters 14 have a relatively small number of coefficients and can therefore adapt quickly.

All the coefficients are constantly being updated by the system, on a sample by sample basis. Thus there is no waiting for a final response before making another adjustment. The sample time is only a small fraction of the fundamental frequency of the primary source 2.

The input signal 16 can be obtained from another moving part of the engine or part of the ignition circuitry; for example.

Figure 2(a) illustrates graphically the response of the reference

signal generator 15 when in the form of a tracking filter. That is to say, a filter having a centre frequency so controlled that the filter output frequencies have a constant ratio to the dominant input frequencies, so that in Figure 2(a), the frequency f_o is N x (engine crankshaft rotation rate), and the reference signal 4 (Figure 1) contains only the first N harmonics of the engine rotation rate, where N is an integer. If the filter input signal is a voltage pulse train as represented by Figure 2(b), where 8T is the periodic time of the engine rotation rate, the first eight harmonics of the engine rotation rate are present in the reference signal 4. The spectrum, by Fourier analysis, of the reference signal 4 is then illustrated by Figure 2(c), in which A is amplitude. With the response of Figure 2(a), only the first six harmonics would be usable.

The tracking filter comprising the signal generator, can be in the form of charge coupled devices having a switching frequency locked to the engine crankshaft rotation rate.

An alternative form of reference signal generator, which employs a plurality of tracking band pass filters is illustrated by Figure 3 in which the input signal 16, a square wave at, for example, 128 times the engine drive shaft rotation rate, is divided first by 32 and then by 2. Division by 32 is achieved by a divider 6 which produces a square wave signal 7 at four times the engine crank-shaft rotation rate, which signal is supplied to a bandpass filter 17. The filter 17 has a centre frequency f_4 which is arranged to track the fundamental frequency of the square wave signal 7 supplied thereto. The further division by 2 is achieved by a divider 8 which produces a square wave signal 9 at twice the engine drive shaft rotation rate. The signal 9 is then supplied to a bandpass filter 18 having a centre frequency f_2 and which is arranged to track the fundamental frequency of the square wave signal 9 supplied thereto. The bandpass filters 17 and 18 produce respectively sinusoidal output signals 7a, 9a at f_2 and f_4 which are linearly summed at an adder 19 to produce the required reference signal 4.

Further dividers and tracking bandpass filters can, of course, be incorporated in the circuit of Figure 3 so that the reference signal 4 contains the desired set of engine order frequencies.

Another form of reference signal generator may comprise a fixed

frequency filtering circuit which selects harmonics and/or subharmonics from an input signal rich in the harmonics of the engine crankshaft rotation rate or firing rate. The filtering circuit may comprise a bandpass filter having a centre frequency fixed at the frequency of a pronounced resonance excited in the enclosure 10 (Figure 1) by the engine 2 or other primary source of vibration. For example, a bandpass filter may be arranged to have a response that models the acoustic response of a motor car passenger compartment to the engine.

A further form of reference signal generator may comprise a plurality of phase lock loops used to generate sinusoidal signals having respective frequencies with integer relationships to a square wave input signal from the engine 2 or other primary vibration source. The sinusoidal signals may then be added together to form the required reference signal. Thus a reference signal 4 comprising specific harmonics and/or harmonics locked to the primary source fundamental, such as engine crankshaft rotation rate, is generated.

An alternative generator for such a reference signal is illustrated in Figure 4 in which a square wave 20 at the primary source fundamental is used to control a plurality of tunable digital oscillators 25, 26 each producing a sinusoidal signal at a chosen harmonic or subharmonic frequency to be added at the adder 19 which produce the reference signal 4 by simple addition of the sinusoidal signals.

In the signal generator 27 of Figure 4, the square wave signal 20, which is at an engine crankshaft rotation rate, is supplied to a bistable circuit 21 which divides the rate signal by two and thereby produces a pulse train signal 20<u>a</u> in which the duration of each pulse is equal to the prevailing periodic time of the square wave signal 20. This periodic time is then measured by a counter 22 which is enabled throughout the duration of each positive pulse from the bistable circuit 21 and counts clock pulses supplied by a clock pulse generator 23. The clock pulses are generated at a fixed, suitably high rate f<u>c</u>.

The contents of the counter 22 are read at the end of each positive pulse from the bistable circuit 21 by a trigonometric function generator 24. The generator 24 is triggered by the trailing pulse of

each positive pulse from the circuit 21 and generates two digital outputs representing respectively $\cos(2\ _0)$ and $\cos(4\ _0)$ where $_0$ is given by

$$\omega_0 = 2\pi f_c/Nf_s$$

in which N is the number of clock pulses counted by the counter 22 in the duration of one positive pulse from the circuit 21, and f_s is a sample rate used in the two digital oscillators 25 and 26 which receive respectively the digital outputs $\cos(2\omega_0)$ and $\cos(4\omega_0)$ from the function generator 24. The digital sinusoidal outputs from the two oscillators 25 and 26 are superposed by a digital adder 19^1 which supplies the reference signal 4 as a digital signal. The trigonometric function generator 24, oscillators 25 and 26 and adder 19^1 can be implemented by a microprocessor with a suitable program. The microprocessor can be used to produce the reference signal 4 in the form :-

$$x_I(n) = \delta(n) + 2\cos(I\omega_0)x_I(n-1) - x_I(n-2)$$

where I is the order of the harmonic or subharmonic being generated, $\delta(n)$ is a unit sample sequence which initiates the simulation of the oscillator, and n is the sample number.

Figure 5 represents in block form an active sound control system 30 for reducing the level of engine generated noise in the passenger compartment of a motor car. The car is provided with an ignition circuit including a low tension coil 31 from which a voltage signal 32 at the firing rate of the engine is taken and supplied to a waveform shaper 33 which in response thereto produces a pulse train at the engine firing rate. It is assumed in the present example that the engine firing rate is twice the engine crankshaft rotation rate f_o. Thus the shaper 33 provides a signal having a fundamental frequency which is a single harmonic, $(2f_o)$, of the crankshaft rate. A reference signal generator is provided in the form of a proprietory tracking filter 34, manufactured by Bruel and Kjaer under type number 1623. The tracking filter 34 receives the output of the shaper 33 as an input signal and as a trigger signal and produces a sinusoidal output signal at the selected harmonic $2f_o$. This sinusoidal signal is sampled with an analog to

digital converter 35 to produce a reference sequence x(n) of digitised samples which are supplied as data to a processor and memory unit 36.

Mounted within the motor car passenger compartment 10 (not shown in Figure 5) are two loudspeakers 37_1 and 37_2, which are in positions normally used for car stereo reproduction. The loudspeakers 37_1, 37_2 are driven by a multiplexer 38 through respective low pass filters 39 and output amplifiers 40. The filters 39 have a cut off frequency of 460 Hz and are provided to prevent aliasing. The multiplexer 38, which contains sample and hold circuits for each output, is controlled by the processor and memory unit 36, through a control line 55, and receives a single input signal 57 from a digital to analog converter 41. The purpose of the loudspeakers 37_1 and 37_2 is to generate, in the passenger compartment, audio waves that will cancel those set up directly by mechanical transmission from the engine to the compartment. The digital to analog converter 41 is supplied by the processor and memory unit 36 with output data 58 which consists of two interleaved sequences of digitised samples $y_1(n)$ and $y_2(n)$. The data 58 is converted by the converter 41 into interleaved sequences of analog samples and separated into respective sequences by the multiplexer 38 for application to the low pass

filters 39. Thus, in effect, the loudspeaker 37_1 is driven by the sequence $y_1(n)$ and the loudspeaker 37_2 is driven by the sequence $y_2(n)$. In Figure 5 each sequence of data 58 is represented by the expression $y_m(n)$, so that in this example m may be 1 or 2.

In order to ensure that the acoustic outputs from the loudspeakers 37_1 and 37_2 have the correct phase and amplitude to effect cancellation of the engine noise, error signals are picked up from the passenger compartment and utilised by the processor and memory unit 36. Acoustic error signals, if present, are sensed by four microphones 42_1, 42_2, 42_3 and 42_4, which are placed respectively either side of a driver headrest and a passenger headrest, there being only two seats in the compartment in the present example. The electrical outputs from the microphones 42_1 etc. are respectively amplified by amplifiers 43 and passed through low pass filters 44 to a four-input multiplexer 45 which supplies a single analog output to an analog to digital converter 46. The filters 44 are provided to prevent aliasing and have a cut-off frequency of 460 Hz.

The multiplexer 45 is controlled by the processor unit 36 by way of control line 56.

The multiplexer 45 and the converter 46 convert the four filtered microphone outputs into a data stream 59 comprising four interleaved sequences of digitised samples $e_1(n)$, $e_2(n)$, $e_3(n)$ and $e_4(n)$, which correspond respectively to the filtered outputs of the

microphones 42_1, 42_2, 42_3 and 42_4. In Figure 5, each sequence is represented by e (n) so that in this example ℓ may be 1, 2, 3 or 4.

The processor and memory unit 36 receives a square wave signal 60 at 1.2 kilohertz from a sample rate oscillator 47 which determines the rate at which the converters 35, 41 and 46 convert samples and the frame duration of processing carried out by the unit 36. Thus in the present example, the unit 36 completes its processing frame within 833 milliseconds. A crystal clock oscillator 61 with a frequency of 10 Megahertz is included in the unit 36.

The unit 36 simulates two adaptive filters, each having two coefficients, so that :-

$$y_m(n) = w_{m0}x(n) + w_{m1}x(n - 1)$$

describes the relation between the output sequence $y_m(n)$ to a loudspeaker and the reference signal x(n), where the coefficients are w_{m0} and w_{m1}. Hence with the two loudspeakers 37_1 and 37_2 :-

$$y_1(n) = w_{10}x(n) + w_{11}x(n - 1) \text{ and}$$

$$y_2(n) = w_{20}x(n) + w_{21}x(n - 1).$$

The values of the coefficients w_{m0} and w_{m1} are calculated by the unit 36 from the relationship :-

$$w_{mi}(n + 1) = w_{mi}(n) + \alpha \sum_{\ell=1}^{4} e_\ell(n) r_{\ell m}(n - i)$$

in which α is a fixed convergence coefficient, $r_{\ell m}(n - i)$ is a value of a filtered reference signal $r_{\ell m}$, and i = 0 or 1.

The filtered reference signal $r_{\ell m}$ is a sequence formed by filtering the reference signal x(n) with a filter that models the effect of the acoustic coupling between the m^{th} loudspeaker and the ℓ^{th} microphone. The unit 36 simulates this filtering as digital FIR (Finite Impulse Response) filtering. Coefficients for the digital FIR filtering are adjusted adaptively during an initialisation program in which a white noise generator 48 is energised.

In the initialisation program, a white noise signal is generated by the generator 48, which is then filtered by a low pass filter 49 to prevent aliasing, the filter 49 having a cut-off frequency of 460 Hz. The signal is subsequently sampled and converted by an analog to digital converter 50. The digital output of the converter 50 is used to drive the loudspeakers 37_1 and 37_2, by way of the processor

and memory unit 36, and the resulting digital input to the unit 36 from the microphones 42_1, 42_2, 42_3 and 42_4 is used to determine the values of reference filter coefficients $c_{\ell mj}$ where $j = 0, \ldots, 34$. The unit 36 performs a 35 coefficient FIR modelling of the impulse response between the m^{th} loudspeaker and the ℓ^{th} microphone at the j^{th} sample. Such modelling is described in "Adaptive Signal Processing" by B. Widrow and S.D. Stearns, published in 1985 by Prentice Hall.

The filtered reference sequence is then given by :-

$$r_{\ell m}(n) = \sum_{j=0}^{34} c_{\ell mj}x(n - j).$$

The operation of the unit 36 is such that, having obtained the error samples $e_\ell(n)$ and the filtered reference signal $r_{\ell m}(n)$, each adaptive filter coefficient w_{mi} for each output $y_m(n)$ is updated by a quantity proportional to the sum of the computed products of $e_\ell(n)$ and $r_{\ell m}(n - i)$ in accordance with the equation :-

$$w_{mi}(n + 1) = w_{mi}(n) + \alpha \sum_{\ell=1}^{4} e_\ell(n) r_{\ell m}(n - i)$$

The new set of adaptive coefficients w_{mi} is then stored and used to filter the next sample of the reference signal, $x(n + 1)$.

The unit 36 includes RAM for temporary storage and computation, and EPROM for program storage. Calculated coefficients w_{mi} and $c_{\ell mj}$,

and reference sequences $r_{\ell m}(n)$ are held in RAM. The convergence coefficient is entered at a set of manually operable switches (not shown).

Preferably the unit 36 includes a Texas Instruments TMS 32010 microprocessor. The input signal rate from the ignition circuit, including the low tension coil 31, is 100 Hz to 200 Hz, and the waveform shaper 33 is a monostable circuit triggered by the leading edge of the input signal to produce pulses of a constant width which is small relative to the sample period set by the sample rate of 1.2 kilohertz. The low pass filters 39, 44 and 49 are active filter modules supplied by Kemo Limited under No. 1431/L.

Only a small amount of separate additional RAM is required with the above-mentioned TMS 32010 microprocessor, which has considerable internal RAM and operates as described in the TMS 32010 Users' Guide published in 1983 by Texas Instruments Inc. Data buses between the unit 36 and the converters 35, 41, 46 and 50 are 12 bit buses. Other buses and lines required for synchronisation and control are omitted for clarity. It will be noted that the values of the coefficients w_{mi} and $c_{\ell mj}$ can be initially set to zero.

A family of algorithms which are alternatives to the SNM algorithm uses a bank of adaptive digital filters working in parallel for each secondary source. Each individual filter is fed by a reference signal containing a subset of the harmonics or subharmonics to be controlled. For example, Figure 6 of the accompanying drawings shows two parallel FIR filters 70 each fed by pure tone reference signals 71, at the second and fourth engine order frequencies in this case. The outputs 72 of these filters are added together by an adder 73 to form an output 74 to the secondary source. Each of the parallel filters 70 may be updated by any of the algorithms discussed above. For example, the said stochastic gradient algorithm may be modified so that :-

$$w_{Imi}(n + 1) = w_{Imi}(n) + \alpha \sum_{\ell=1}^{L} e_{\ell}(n)\, r_{\ell mI}(n - 1)$$

where w_{Imi} is the i'th coefficient of the FIR filter fed from the I'th harmonic of the engine, driving the m'th secondary source.

The advantage of such an algorithm is that each harmonic frequency is controlled independently and the convergence of an harmonic does not couple with the convergence of any other harmonic, as is the case when a 2I coefficient filter is used to filter I harmonics simultaneously. In this case I filters each with 2

coefficients could be used to filter I harmonics individually, and their response combined afterwards. The disadvantage of this algorithm is that a filtered reference signal needs to be generated for each source (m), sensor (ℓ) <u>and</u> harmonic (I), to give each $r_{\ell mI}(n)$.

Another approach to controlling a number of harmonics is to take the Fourier transform of each of the error signals and to update a set of coefficients controlling each harmonic of each secondary output independently. The outputs of each of these filters are then combined together, for each secondary source, via an inverse Fourier transform, to generate the output waveform for this source, as indicated in Figure 7.

For a single harmonic in the frequency domain, the complex value of the 'th error signal will be given by

$$E_\ell = A_\ell + \sum_{m=1}^{M} c_{\ell m} w_m$$

where A_ℓ is the value of E_ℓ with no active control, w_m is the complex amplitude of the voltage to the m'th secondary source and $c_{\ell m}$ is the complex transfer function between the ℓ'th sensor and m'th source at the frequency of the harmonic of interest.

This may be expressed in matrix form as :-

$$\underline{E} = \underline{A} + \underline{CW}$$

where :-

$$\underline{E}^T = (E_1 \quad E_2 \quad \ldots \quad E_L)$$
$$\underline{A}^T = (A_1 \quad A_2 \quad \ldots \quad A_L)$$
$$\underline{W}^T = (W_1 \quad W_2 \quad \ldots \quad W_M)$$
$$\underline{C} = \begin{pmatrix} C_{11} & C_{12} & \cdots & C_{1M} \\ C_{21} & C_{22} & \cdots & \cdot\cdot \\ \vdots & & & \\ C_{L1} & \cdots\cdots & & C_{LM} \end{pmatrix}$$

The cost function in this case may be written as $J = \underline{E}^H\underline{E}$ where the superscript H denotes the complex conjugate of the transposed vector or matrix. Therefore :-

$$J = \underline{A}^H\underline{A} + \underline{W}^H\underline{C}^H\underline{A} + \underline{A}^H\underline{C}\,\underline{W} + \underline{W}^H\underline{C}^H\underline{W}$$

Note that

$$\therefore \qquad \frac{\partial J}{\partial \underline{W}} = 2\underline{C}^H\underline{A} + 2\underline{C}^H\underline{C}\,\underline{W} = 2\underline{C}^H\underline{E}$$

so that the steepest descent algorithm may be written :-

$$\underline{W}_{k+1} = \underline{W}_k - 2\mu\underline{C}^H\underline{E}_k$$

where $\underline{W}_k$ and $\underline{E}_k$ are the complex values of the filter response and error output respectively at the k'th iteration.

This algorithm does not seem to appear in the active control literature. This is probably because the Newton's method algorithm,

below, is no more difficult to implement after initialisation to form the matrix needed to premultiply $\underline{E}_k$. However, in some cases it may be that the matrix $\underline{C}$ changes with time and a separate "identification" algorithm is used in parallel with the adaptive control algorithm to track these changes. In such cases the steepest descent algorithm may be considerably more computationally efficient to implement than that below.

The frequency demain version of the Newton's method algorithm may be written :-

$$\underline{W}_{k+1} = \underline{W}_k - 2\mu(\underline{C}^H\underline{C})^{-1}\underline{C}^H\underline{E}_k$$

A special case of this algorithm does appear in the literature, in which the number of error sensors is equal to the number of secondary sources (L = M), so that $\underline{C}$ is a square matrix, and the algorithm reduces to :-

$$\underline{W}_{k+1} = \underline{W}_k - 2\ \underline{C}^{-1}\underline{E}_k$$

This algorithm has been presented by Pierce (1985, David W. Taylor Naval Ship Research and Development Center Report No. 85/047. An algorithm for active adaptive control of periodic interface). If the convergence coefficient, μ, is set equal to one half, the algorithm also reduces to the iterative matrix algorithm described in White and Cooper (1984, Applied Acoustics 17, 99-109. "An adaptive

controller for multivariable active control"). See also U.K. Patent Specification No. 2,122,052A.

In practice the most advantageous algorithm will probably be derived from a judicious mixture of time and frequency domain concepts. For example, one implementation of a Fourier transformer operating on each error sequence $e_{\ell}(n)$ to produce the inphase and quadrature frequency components of $e_{\ell}(n)$ at $I\omega_0$ is illustrated in Figure 8 wherein integrators 80 and multipliers 81 are used. The slowly varying outputs of this circuit represent the real and imaginary parts of the frequency domain signal E_{ℓ} defined above, so these signals can be used with any of the frequency domain algorithms discussed above to update a bank of adaptive filters driven by this frequency component driving each secondary source, as in Figure 6.

Figure 9 illustrates application of the invention to non-noise, i.e. mechanical, vibration control.

The example illustrated by Figure 9 comprises a modification of the Figure 1 arrangement, wherein the microphone sensors have been replaced by accelerometers 90 and the loudspeakers by mechanical vibrators 91. The accelerometers 90 and vibrators 91 are mounted on surface portions of the enclosure 10.

In another modification, illustrated by Figure 10, a sensing combination of microphones 12 and accelerometers 90 is used, and/or a source combination of loudspeakers 11 and vibrators 91.

A combination of Figure 10 need not be coincident; the components thereof could instead be spaced from each other.

CLAIMS

1. An active vibration control system for reducing vibration generated by a primary source is characterised in that at least one reference signal (4) containing at least one selected harmonics of the said primary source (2) of vibration is supplied to means driving a plurality of secondary vibration sources (11), such that the vibration energy detected by sensor means (12) operable to sense the vibration field established by the primary and secondary sources is reduced.

2. A system according to Claim 1, characterised in that the reference signal (4) is generated by filtering a periodic input signal (16) having its fundamental frequency (9) locked to a predominant frequency (f_2) of the exterior source.

3. A system according to Claim 2, characterised in that filtering is effected by at least one tracking filter (17, 18).

4. A system according to Claim 2, characterised in that the reference signal (4) is generated by at least one tunable oscillator (25) whose frequency is controlled by a signal (20), the primary source fundamental.

5. A system according to any preceding claim, characterised in that the means (36) driving the secondary sources (37, 32_2) comprise a digital processor with data and program memory.

6. A system as claimed in any preceding claim, characterised in that means (70) are provided whereby a plurality of reference signals (71), each containing a single harmonic, are filtered by a bank of filters (70)

which are adjustable independently and whose outputs (72) are combined (74) to form an output to the secondary sources.

7. A system according to any preceding claim, characterised in that it is adapted to operate in accordance with an algorithm which adjusts the outputs from the secondary sources so as to substantially minimise a cost function on a time scale comparable with the delays associated with the propagation of vibration from the secondary sources (11) to the sensor means (12).

8. A system as claimed in Claim 7, characterised in that the algorithm is of the form :-

$$\underline{w}(n+1) = \underline{w}(n) - 2\mu \underline{R}^{T}(n)\, \underline{e}(n).$$

9. A system as claimed in Claim 7, characterised in that the algorithm is of the form :-

$$\underline{w}_{n+1} = \underline{w}_{n} - 2\mu \underline{Q}^{T}(n)\, \underline{e}(n).$$

10. A system as claimed in Claim 7, characterised in that the algorithm is of the form :-

$$\underline{W}_{k+1} = \underline{W}_{k} - 2\mu \underline{C}^{H} \underline{E}_{k}$$

11. A system as claimed in Claim 7, characterised in that the algorithm is of the form :-

$$\underline{W}_{k+1} = \underline{W}_{k} - 2\mu (\underline{C}^{H}\underline{C})^{-1} \underline{C}^{H} \underline{E}_{k}.$$

12. A system according to any preceding claim, characterised in that it makes use of a plurality of closed loops (12, 13, 11), each comprising one of the secondary vibration sources (11), signal adaptive control means (14) and one of the sensors (12).

13. A system according to any preceding claim, <u>characterised in that</u> the vibration controlled comprises sound or noise.

14. A motor car provided with an active vibration control system as claimed in any one of Claims 1 to 13.

15. A system for reducing vibration, substantially as hereinbefore described with reference to the accompanying drawings.

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FIG.1.

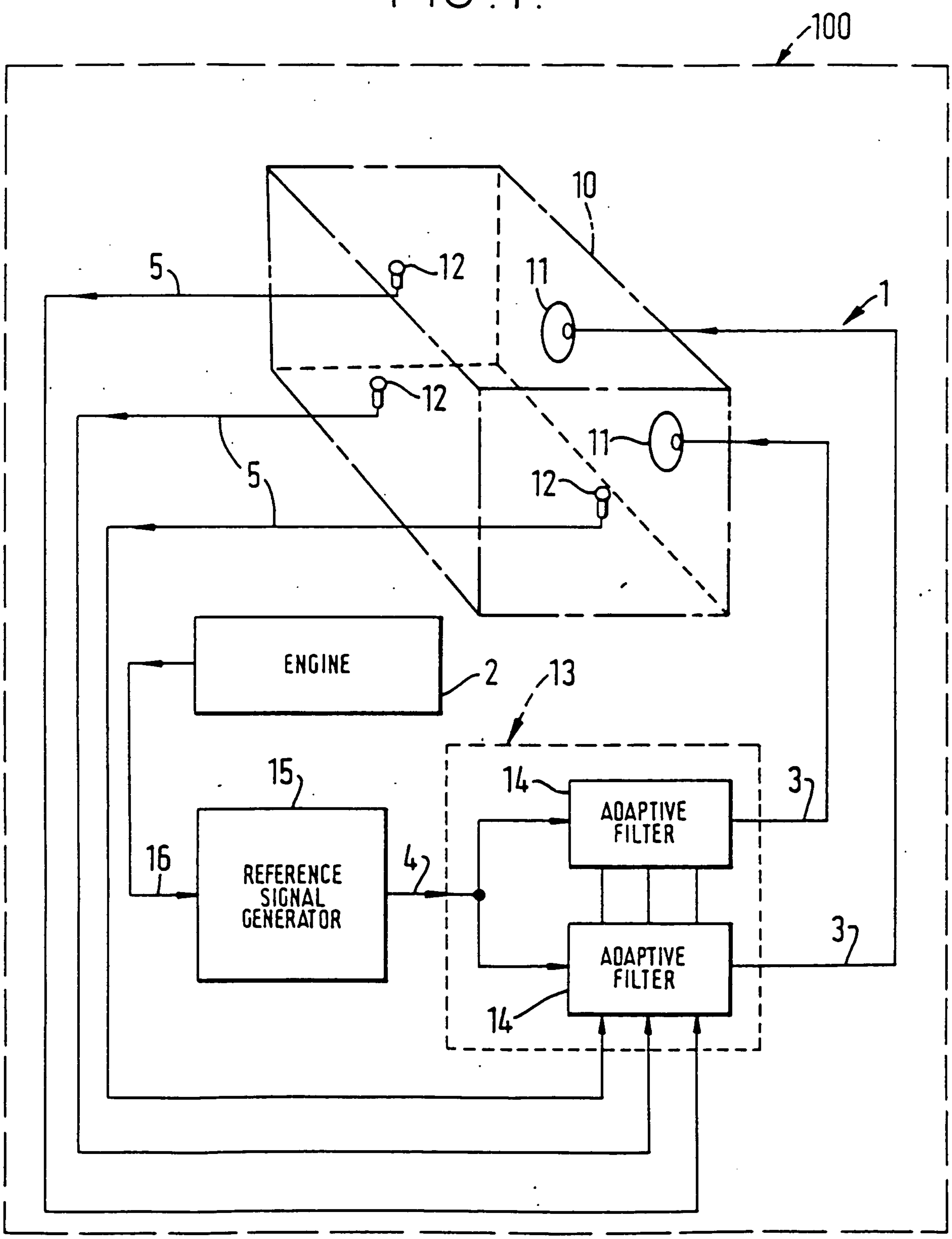

100
10
5
12
11
1
12
11
5
12
ENGINE
2
13
15
14
ADAPTIVE FILTER
3
16
REFERENCE SIGNAL GENERATOR
4
ADAPTIVE FILTER
3
14

FIG. 2.

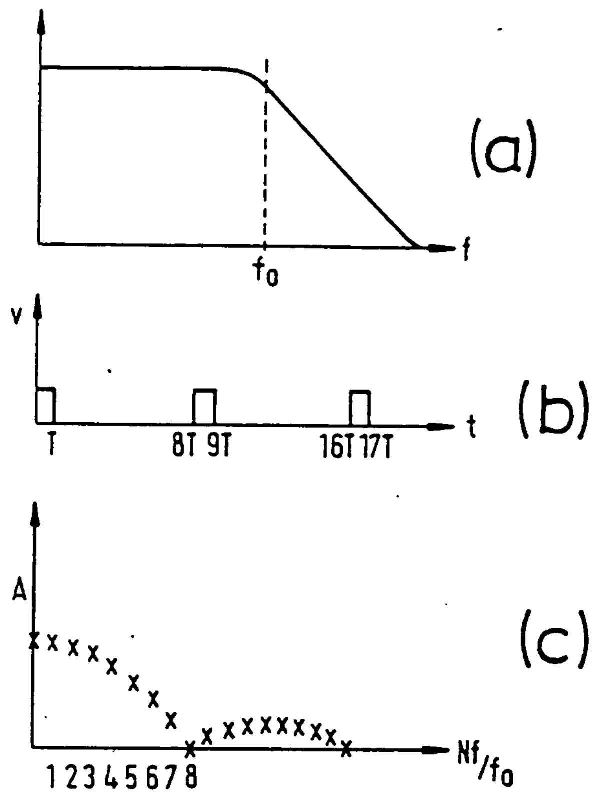
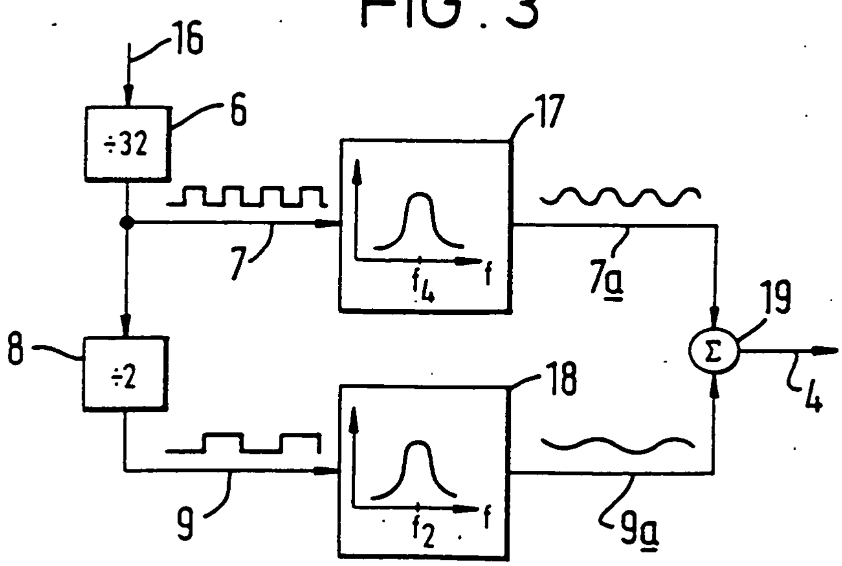
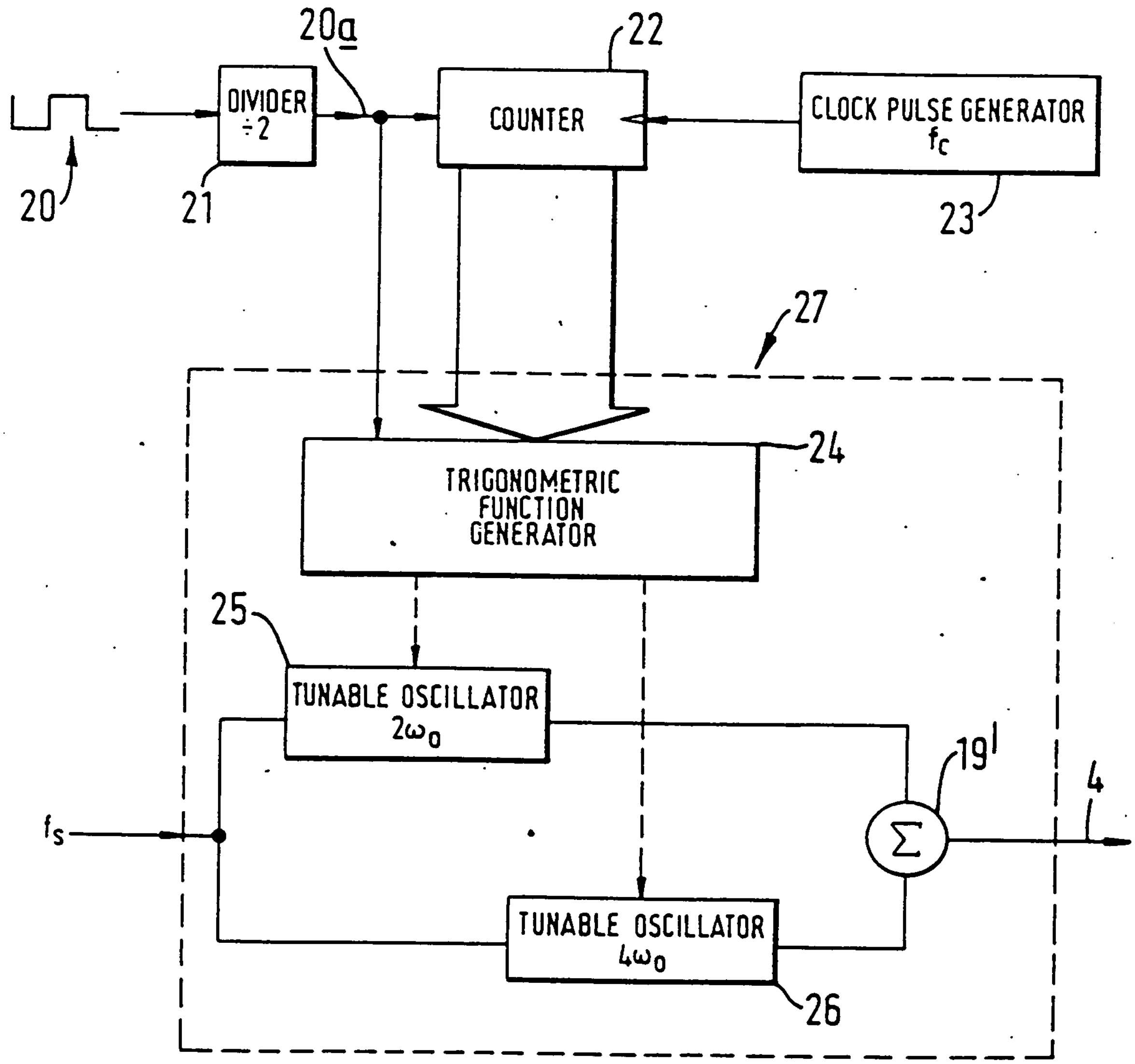


FIG. 3



SUBSTITUTE SHEET

FIG. 4.


20a
22
DIVIDER ÷2
COUNTER
CLOCK PULSE GENERATOR f_c
20
21
23
27
24
TRIGONOMETRIC FUNCTION GENERATOR
25
TUNABLE OSCILLATOR $2\omega_0$
19'
4
f_s
Σ
TUNABLE OSCILLATOR $4\omega_0$
26

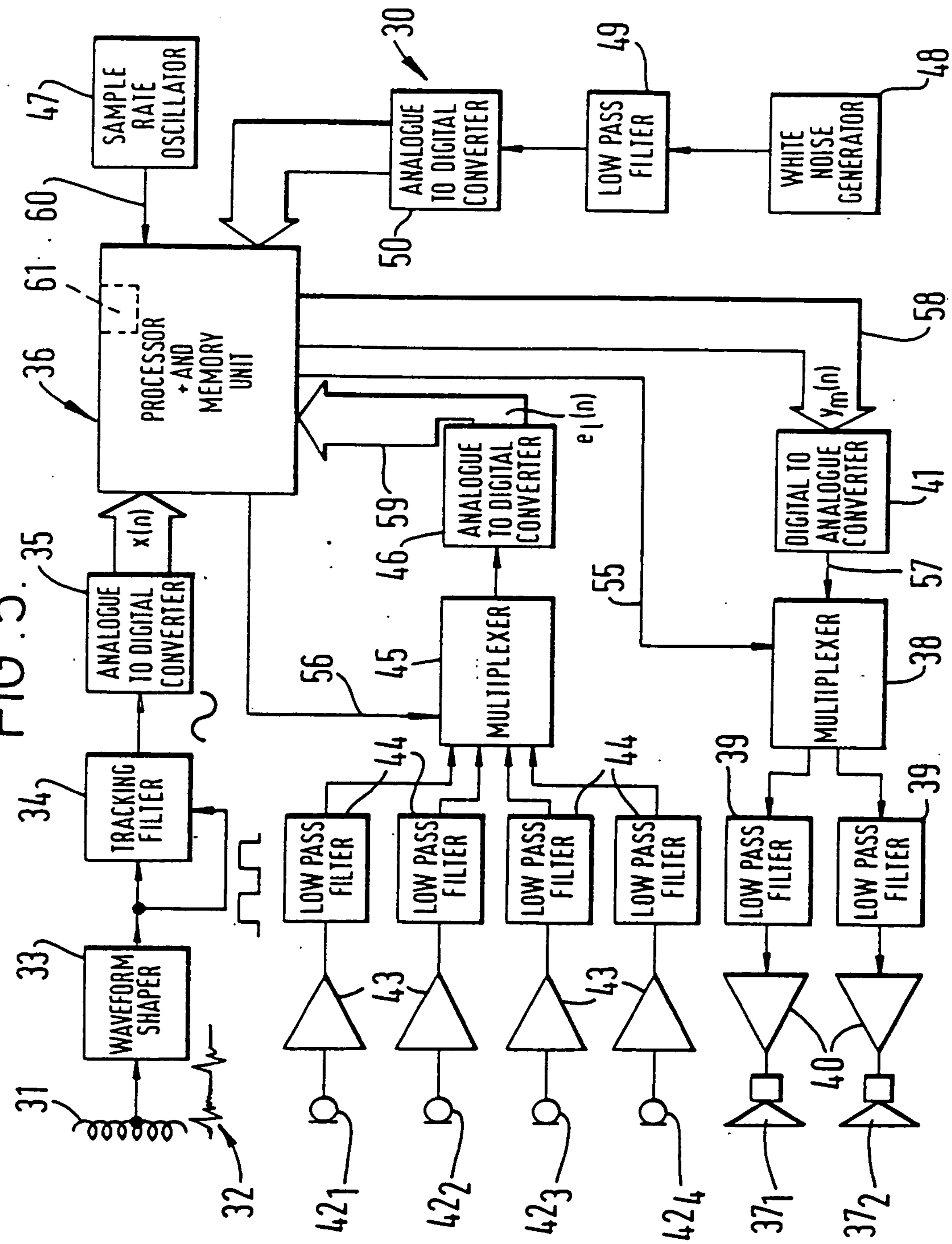
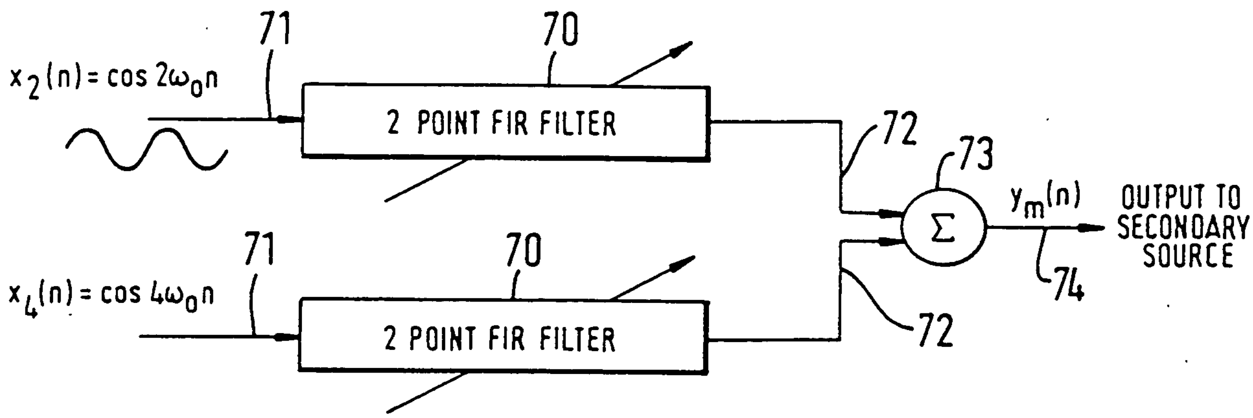

SAMPLE RATE OSCILLATOR
ANALOGUE TO DIGITAL CONVERTER
LOW PASS FILTER
WHITE NOISE GENERATOR
PROCESSOR + AND MEMORY UNIT
$e_l(n)$
$y_m(n)$
DIGITAL TO ANALOGUE CONVERTER
x(n)
MULTIPLEXER
TRACKING FILTER
WAVEFORM SHAPER
LOW PASS FILTER
30
47
49
48
60
61
36
50
58
59
46
41
35
55
57
56
45
38
34
33
31
32
44
43
39
40
42_1
42_2
42_3
42_4
37_1
37_2

FIG. 5

FIG. 6.


$x_2(n) = \cos 2\omega_0 n$
71
70
2 POINT FIR FILTER
$x_4(n) = \cos 4\omega_0 n$
71
70
2 POINT FIR FILTER
72
73
72
Σ
$y_m(n)$
74
OUTPUT TO SECONDARY SOURCE

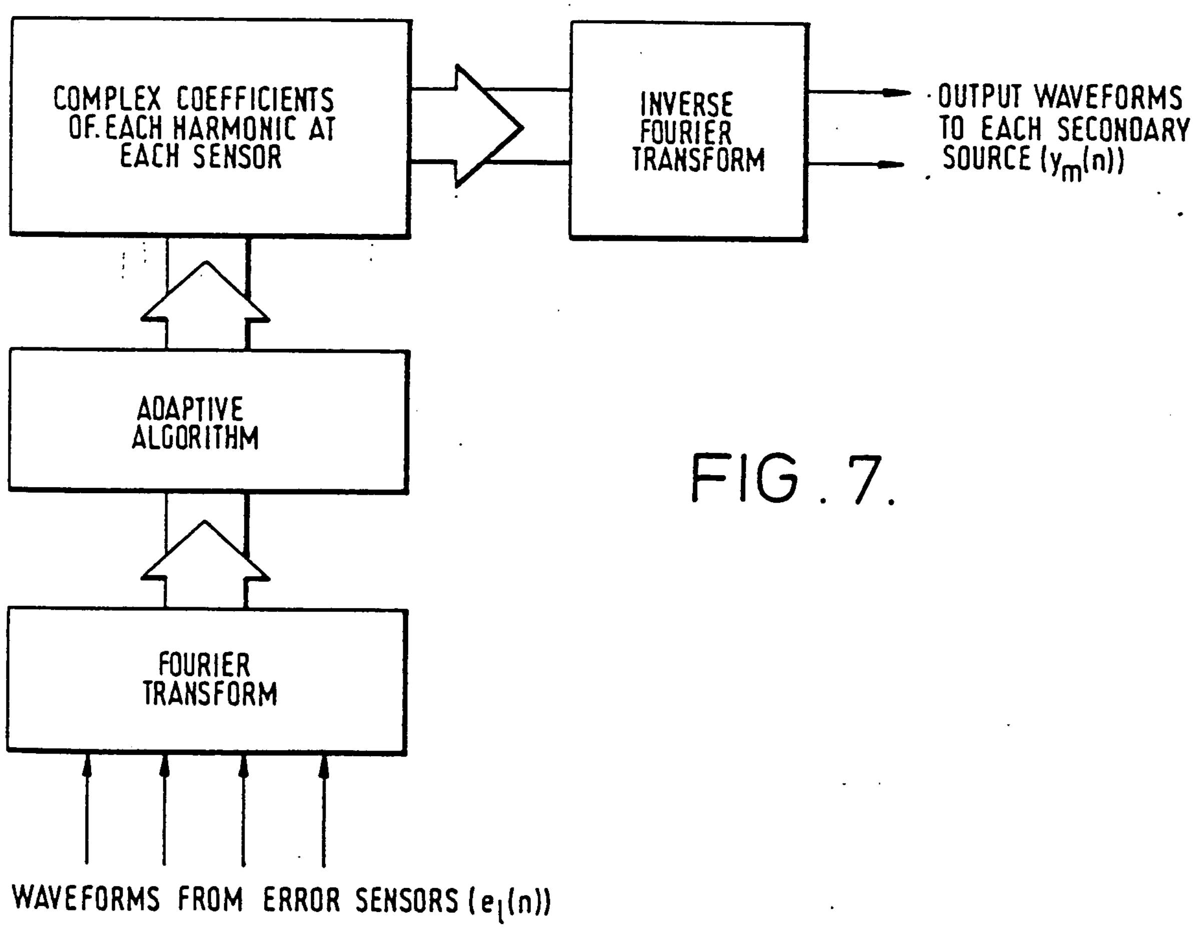

COMPLEX COEFFICIENTS OF EACH HARMONIC AT EACH SENSOR
INVERSE FOURIER TRANSFORM
OUTPUT WAVEFORMS TO EACH SECONDARY SOURCE ($y_m(n)$)
ADAPTIVE ALGORITHM
FOURIER TRANSFORM
WAVEFORMS FROM ERROR SENSORS ($e_l(n)$)

FIG. 7.

FIG. 8.

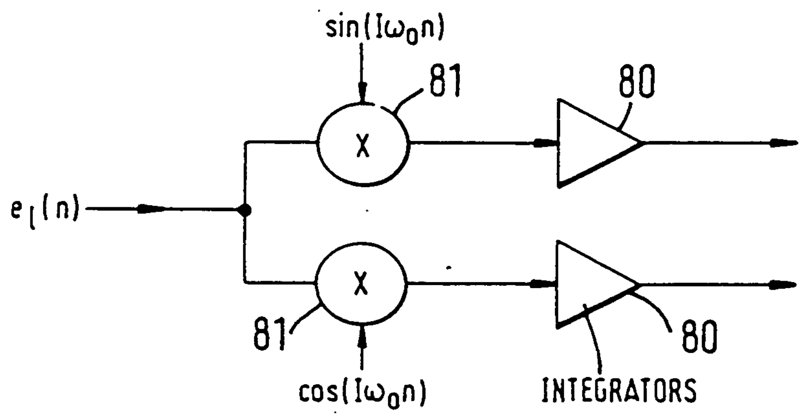
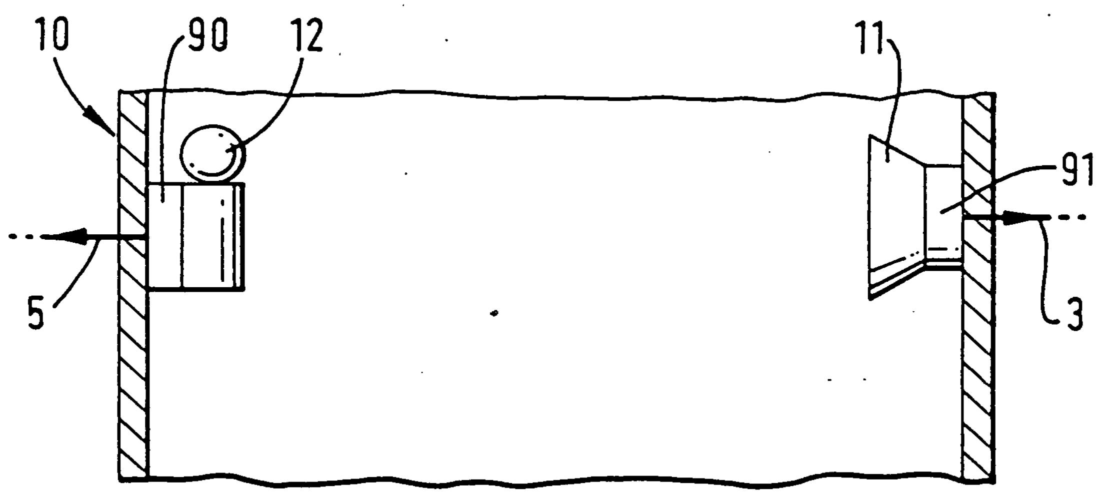
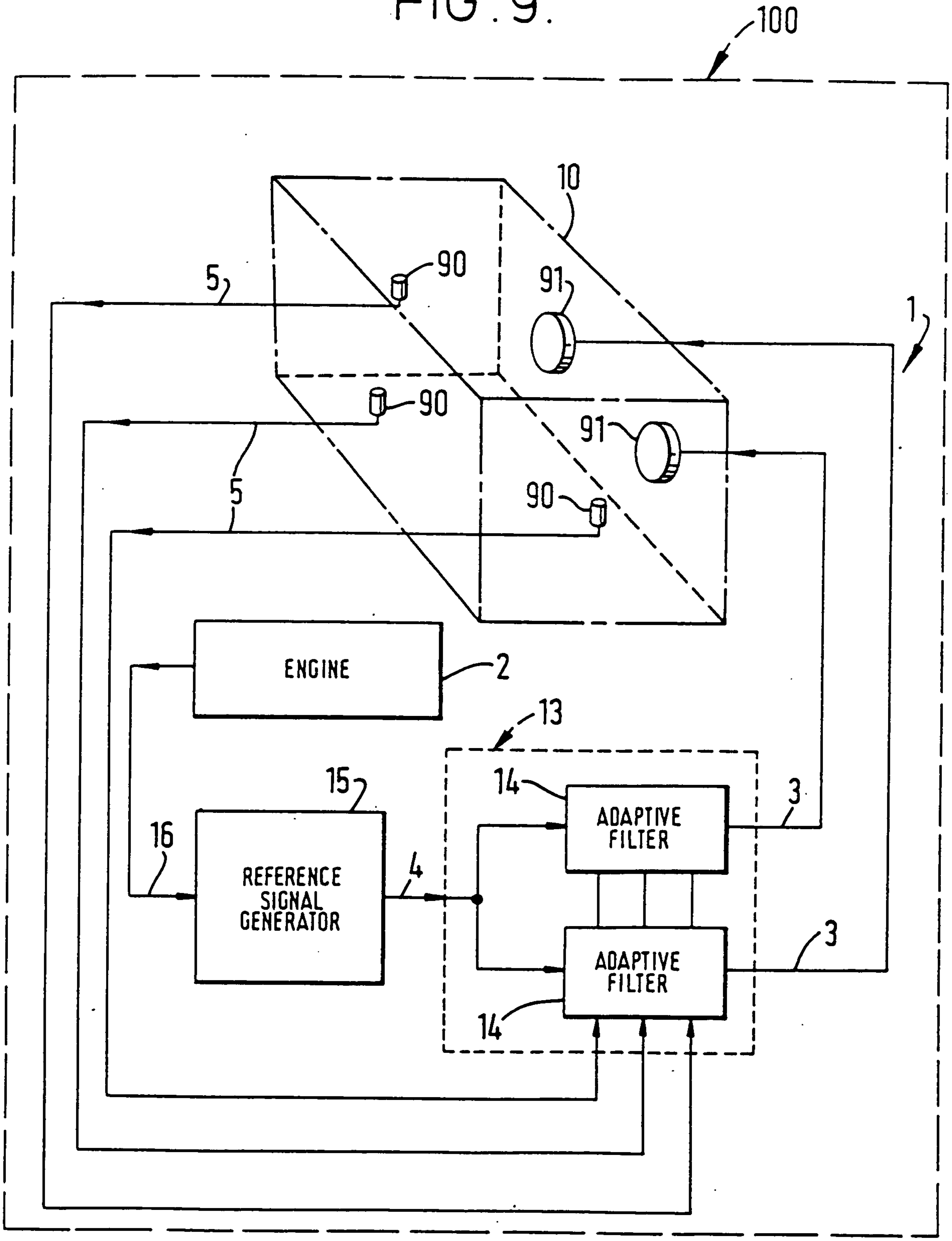

sin($I\omega_0 n$)
81
80
$e_l(n)$
X
X
81
80
cos($I\omega_0 n$)
INTEGRATORS

FIG. 10.


10
90
12
11
91
5
3

7/7

FIG. 9.



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INTERNATIONAL SEARCH REPORT

International Application No PCT/GB 87/00706

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all) [6]

According to International Patent Classification (IPC) or to both National Classification and IPC

IPC^4: G 10 K 11/16

II. FIELDS SEARCHED

Minimum Documentation Searched [7]

Classification System	Classification Symbols
IPC^4	G 10 K

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched [8]

III. DOCUMENTS CONSIDERED TO BE RELEVANT [9]

Category *	Citation of Document, [11] with indication, where appropriate, of the relevant passages [12]	Relevant to Claim No. [13]
A	US, A, 4232381 (RENNICK et al.) 4 November 1980 see column 1, line 49 - column 2, line 7 --	1
A	GB, A, 2149614 (SECRETARY OF STATE FOR DEFENCE) 12 June 1985 see page 2, lines 18-62; figure 2 cited in the application --	1
A	GB, A, 782794 (GENERAL ELECTRIC CO.) 11 September 1957 see page 2, line 96 - page 3, line 2; figure 2 --	1
A	FR, A, 2351466 (SOUND ATTENUATORS LTD) 9 December 1977 see claim 1 cited in the application & GB, A, 1577322 --	1
A	GB, A, 2122052 (THE PLESSEY CO.) 4 January 1984 ./.	1

* Special categories of cited documents: [10]

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

IV. CERTIFICATION

Date of the Actual Completion of the International Search	Date of Mailing of this International Search Report
8th January 1988	12 FEB 1988
International Searching Authority	**Signature of Authorized Officer**
EUROPEAN PATENT OFFICE	P.C.G. VAN DER PUTTEN

Form PCT/ISA/210 (second sheet) (January 1985)

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)

Category *	Citation of Document, with indication, where appropriate, of the relevant passages	Relevant to Claim No
	see abstract cited in the application --------	

4/~ ③

ANNEX TO THE INTERNATIONAL SEARCH REPORT ON INTERNATIONAL PATENT APPLICATION NO. GB 8700706

SA 18992

This annex lists the patent family members relating to the patent documents cited in the above-mentioned international search report.
The members are as contained in the European Patent Office EDP file on 29/01/88
The European Patent Office is in no way liable for these particulars which are merely given for the purpose of information.

Patent document cited in search report	Publication date	Patent family member(s)		Publication date
US-A- 4232381	04-11-80	None		
GB-A- 2149614	12-06-85	None		
GB-A- 782794		None		
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For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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